METHOD AND DEVICE FOR AUTOMATIC CONTROL OF THE FREQUENCY OF A LOCAL OSCILLATOR IN A DS-CDMA TYPE RECEIVER

Field of the Invention

The present invention relates to wireless communication systems, and in particular, to a wireless communication system operating according to the CDMA

5   standard and one that includes a DS-CDMA type receiver with a local oscillator having automatic frequency control.

Background of the Invention

In a wireless communication system, a base

10   station communicates with several remote terminals such as mobile cell phones. Frequency division multiple access (FDMA) and time division multiple access (TDMA) are traditional multiple access schemes used to supply simultaneous services to a number of terminals. The

15   basic idea in FDMA and TDMA systems is to respectively share the available resource at several frequencies or several time intervals so that several terminals can operate simultaneously without causing interference.

Telephones operating according to the GSM

20   standard belong to FDMA and TDMA systems in the sense that transmission and reception takes place at different frequencies, and also at different time intervals. Unlike these systems using frequency

division or time division, CDMA (code division multiple access) systems enable several users to share a common frequency and a common time channel by using a coded modulation. CDMA systems include the CDMA 2000 system,

5   the WCDMA (Wide Band CDMA) system and the IS-95 standard.

A scrambling code is used in CDMA systems associated with each base station to distinguish one base station from another, as readily understood by one

10   skilled in the art. Furthermore, an orthogonal code known as the OVSF code is allocated to each remote terminal such as a mobile cell phone, for example. All OVSF codes are orthogonal with each other so that one remote terminal can be distinguished from another.

15   Before a signal can be sent on the transmission channel to a remote terminal, the signal is scrambled and spread by the base station using the scrambling code for the base station and the OVSF code for the remote terminal. Consequently, the signal

20   symbols are transformed into chips. The chip rate is higher than the symbol rate.

DS-CDMA systems refer to CDMA systems using spread spectrum signals. Conventionally, this type of receiver includes a radio frequency analog stage

25   connected to an antenna to receive a spread spectrum signal.

The radio frequency stage comprises a low noise amplifier and two processing channels comprising mixers, filters and conventional amplifiers. The two

30   mixers receive two corresponding signals at a phase difference of 90° from a phase locking loop comprising a local oscillator. After the frequency has been transposed in the mixers, the two processing channels

define two paths I and Q, which are in quadrature. The signal is then transposed in the frequency base band of the signal.

5 After digital conversion in analog/digital converters, the I and Q paths are output to a reception processing stage that conventionally comprises a Rake receiver. The Rake receiver is typically used in mobile cell phones operating in a DS-CDMA communication system, and is used for time alignment, descrambling,
10 despreading and a combination of delayed versions of the initial signals so as to output information contained in the initial signals.

The frequency transposition of the signal received in the base band usually leads to
15 inaccuracies. This is mainly due to the quality of the local oscillator which then causes a residual frequency offset error. The digital processing stage of the receivers usually comprises an automatic frequency control (AFC) algorithm, the objective of which is to
20 minimize the residual frequency error to bring it down to an acceptable value that has no influence on subsequent signal processing.

There are several available methods of frequency correction. Automatic frequency control
25 techniques used in narrow band communications may be used for frequency control after despreading. Several methods are discussed in the following references: A. N. D'Andrea and U. Mengali, "Design Of Quadricorrelators For Automatic Frequency Control
30 Systems", IEEE Trans. Commun., vol. 41, pp. 988-997, January 1993; M. L. Fowler and J.A. Johnson, "Extending The Threshold And Frequency Range For Phase-Based Frequency Estimation", IEEE Trans. Signal Processing,

vol. 47, pp. 2857-2863, October 1999; and A. Wannasarnmaytha, S. Hara and N. Morinaga, "Two Step Kalman Filter Based AFC For Direct Conversation-Type Receiver In LEO Satellite Communications", IEEE Trans.

5   Commun., vol. 49, pp. 246-253, January 2000.

However, performances of these techniques do not satisfy a number of requirements of third generation receivers designed to operate under the CDMA standard. For example, a wide correction frequency

10   range, fast correction and low implementation complexity are not satisfied. More precisely, correction ranges of automatic frequency control techniques used in a narrow band are insufficient when the frequency is high and the frequency offset is of

15   the order of the symbol rate, as in third generation wireless communication systems for example.

Automatic frequency control techniques with dual filters and with balanced quadricorrelators as disclosed in the D'Andrea et al. article use correction

20   ranges on the order of the symbol rate. However, these techniques involve the use of complicated filters to eliminate inductive noise and require precise knowledge about the transmission channel.

Other automatic frequency control techniques

25   (Jing Lei and Tung-Sang Ng, "New AFC Algorithm For A Fully-Digital MDPSK DS/CDMA Receiver", ISCAS 2001, vol. 4, pp. 294-297) also use correction ranges on the order of the symbol rate, but their implementation is very complex due to the fact that they must be applied

30   before signal despreading.

Summary of the Invention

In view of the foregoing background, an object of the present invention is to provide automatic frequency control for a local oscillator of a DS-CDMA
5   type receiver with a low manufacturing complexity, and operating at the symbol level with a correction range on the order of the symbol rate which is approximately twice the correction range of automatic narrow band control techniques.

10   Another object of the present invention is to provide automatic frequency control that has a much faster and much more precise convergence than existing automatic frequency controls.

These and other objects, advantages and
15   features in accordance with the present invention are provided by a process for automatic frequency control of a local oscillator of a DS-CDMA type receiver in which a known spread spectrum signal is received (for example, a pilot signal) that is transformed into a
20   sampled digital signal formed from symbols and with a despreaded frequency spectrum. A residual frequency error is determined for each symbol. This error is filtered and the local oscillator frequency is corrected with this filtered error.

25   According to a general characteristic of the invention, the first determined residual frequency error f_{e1} is memorized, and the average of the absolute values of a predetermined number of successive residual frequency errors is determined. This average is
30   compared with a predetermined threshold. If this average is greater than the threshold, the local oscillator frequency is corrected with an error equal to $-sgn(f_{e1})(1/T - |f_{e1}|)$, where sgn is the sign

function, | | is the absolute value function and T is
the duration of a symbol before determining the next
residual frequency error associated with the next
symbol.

5   Since residual frequency errors are
advantageously filtered digitally, if the average is
greater than the threshold, the memory of the digital
filter is set to zero before the next residual
frequency error is filtered.

10   Residual frequency errors are advantageously
filtered with a first order matching filter. More
precisely, when the estimated residual frequency error
is large, this estimated error is not filtered at the
limit. This causes fast convergence towards the

15 frequency offset value, but obviously once this value
has been reached, the oscillations around this value
are large. The pass band of the matching filter also
reduces when the frequency error reduces.

        Thus, according to one embodiment, the

20 current correction applied to the local oscillator is
equal to (1-b) times the previous correction plus b
times the current residual frequency error, and the
coefficient b is chosen to be close to 1 if the current
residual frequency error is greater than a first

25 predetermined limiting value, whereas the coefficient b
is chosen to be close to zero if the current residual
frequency error is less than a second predetermined
limiting value. For example, the first predetermined
limiting value may be equal to the symbol rate divided

30 by 7, for example.

        The first and the second limiting values may
be identical or they may be different. If they are
different, several intermediate limiting values may be

used between these two values, and they may be stored in a table so that the value of b can be reduced in steps, for example.

Another aspect of the present invention is
5 directed to an automatic device for controlling the frequency of a local oscillator in a DS-CDMA type receiver comprising one input for receiving a known spread spectrum signal, and pre-processing means for transforming this signal into a sampled digital signal
10 formed from symbols with a despreaded frequency spectrum. First calculation means determine a residual frequency error for each symbol. Filter means filters this error, and correction means corrects the frequency of the local oscillator with this filtered error.

15 According to one general characteristic of the invention, the device also comprises a memory to store the first determined residual frequency error f_{e1}, and second calculation means for determining the average of the absolute values of a predetermined
20 number of successive residual frequency errors. Comparison means compares this average with a predetermined threshold, and if this average is greater than the threshold, the correction means corrects the frequency of the local oscillator with an error equal
25 to $-sgn(f_{e1})(1/T - |f_{e1}|)$, where sgn is the sign function, $|\ \ |$ is the absolute value function, and T is the duration of a symbol before the first calculation means determines the next residual frequency error associated with the next symbol.

30 According to one embodiment, the filter means comprises digital filtering, and the device comprises control means that, if the average is greater than a

threshold, will reset the memory of the digital filter before filtering the next residual frequency error.

Yet another aspect of the present invention is directed to a DS-CDMA type receiver comprising an automatic frequency control device as defined above. This type of receiver may be a mobile cell phone, for example.

Brief Description of the Drawings

Other advantages and specific features of the invention will become clear after examining the detailed description of embodiments and uses which are in no way limiting, and the appended drawings in which:

Figure 1 is a block diagram of a device in accordance with the present invention; and

Figure 2 is a graph of two curves illustrating one of the problems addressed by the present invention.

Detailed Description of the Preferred Embodiments

In Figure 1, the reference TP denotes a mobile cell phone belonging to a wireless DS-CDMA type communication system. Conventionally, the analog reception stage of the mobile cell phone TP comprises a mixer MX receiving the incident signal from the antenna. The mixer is preceded and followed by amplifiers, which are not shown in the figures simplify the drawing. Similarly, only one of the processing channels is shown in Figure 1, which is processing channel 1.

The frequency transposition signal is derived from a phase locking loop comprising a local oscillator that is voltage controlled and is referenced VCO. After

transposition of frequencies in the mixer MX and
analog-digital conversion in an analog-digital
converter CAN, the signal is a digital base band signal
sampled at a sampling rate much higher than the symbol
5   rate equal to 1/T (where T is the duration of a
symbol). This digital signal is then output to a Rake
receiver RR that in particular performs frequency
despreading of the signal.

Thus, if it is considered that the mixer
10   input signal is the pilot signal sequence, the signal
y(k) may be modeled after despreading the pilot
sequence using the following formula:

$$y(k) = \beta b(k) A e^{j(\varphi(k)+\phi(k)+2\pi f_e(k))} + \eta(k) \tag{I}$$

In this formula: $Ae^{j\varphi(k)}$ is the sequence transmitted to
15   the pilot, where A is a real; $e^{j\varphi(k)}$ is the phase
distortion of the transmission channel; b(k) is the
amplitude distortion of the channel; f_e is the residual

frequency error; and $\eta(k)$ is complex noise and β is a
real attenuation coefficient between 0 and 1 that
20   decreases as the residual frequency error f_e increases.

The digital stage of the mobile cell phone TP
is provided with an automatic frequency control device
DAFC, which is essentially performed via software
within a processor. The role of this device DAFC is to
25   respectively estimate successive frequency errors f_e
associated with successive symbols, and after digital-
to-analog conversion in a CNA converter, to control the
oscillator VCO to minimize (i.e., reduce) the residual
frequency error. More precisely, the oscillator
30   voltage is controlled so as to control its frequency,

i.e., the frequency of the transposition signal output by this oscillator.

The device DAFC comprises first calculation means MC1 performed by software and is capable of
5 determining a residual frequency error f_e for each symbol. These successive frequency errors are determined conventionally by an autocorrelation of two consecutive symbols. This is well known to one skilled in the art, and therefore, will not be discussed in
10 further detail.

In general, and unless mentioned otherwise below, each residual frequency error is filtered by a low pass digital filter FLT. Correction means includes a digital-to-analog converter CNA for correcting the
15 local oscillator frequency with the filtered error.

Several cases can then arise. When the first determined residual frequency error is less than half of the symbol rate, in other words less than 1/2T, the corrections made by the device DAFC are good and the
20 absolute values of successive residual frequency errors can converge towards 0. This is not the case when the first determined residual frequency error is more than 1/2T. However, the invention will overcome this problem as will be described in more detail below.

25 If we consider a pilot signal sequence despread in the frequency domain, it can be seen that since the pilot sequence has a constant value and the despreading process requires an averaging and under sampling operation, the complete process can be seen as
30 a convolution of the pilot sequence in the time domain with a rectangle function over a time length equal to the duration of a symbol T, followed by under sampling from the chip rate towards the symbol rate 1/T.

In the frequency domain, this process leads to a spectrum resulting in the product (due to the constant value of the pilot sequence) with a $(\sin x)/x$ function in which the width of the main lobe is equal
5 to $1/T$, then all replicas of this spectrum with a frequency offset of k/T are summated together (for each integer value of k).

The spectral density obtained after despreading is not modified when there is a frequency
10 shift error, but the Dirac corresponding to the pilot sequence is offset from the origin by an offset corresponding to the error. When the residual frequency error is less than $1/2T$, there is no spectrum folding phenomenon, and in this case successive errors will
15 converge towards zero as described above. However, when the absolute value of the residual frequency error is more than $1/2T$, there is a spectrum folding phenomenon.

For example, suppose that there was frequency offset equal to $-1/2T-\Delta f$, where $0 < \Delta f < 1/2T$. Due to
20 the folding, this frequency error will be interpreted incorrectly as being equal to $1/2T-\Delta f$. This is illustrated in Figure 2, which shows superposition of the Dirac D1 corresponding to the frequency error equal to $1/2T-\Delta f$, and the Dirac D2 corresponding to the
25 frequency error equal to $-1/2T-\Delta f$.

It can be seen in this example that there is also a large difference in the power between the folded Dirac D2 and the unfolded Dirac D1. With the unfolded signal, the estimate of the frequency error will be
30 correct and its compensation will shift Dirac D1 towards the left towards a zero frequency error along curve C1, and by increasing its energy.

However, compensating the badly interpreted frequency error of the folded signal will also offset the Dirac D2 towards the left towards the origin, but the frequency error will increase at the same time. In
5　this case, the already low energy of the signal will decrease towards zero along curve C2.

The decrease in the signal-to-noise ratio of the folded signal will be so high that the estimate of the frequency error will be very bad, which will cause
10　large variations in the estimates of successive frequency errors. According to the invention, this large variation of residual frequency errors will be used to detect and correct frequency errors larger than the Nyquist frequency, in other words greater than
15　1/2T.

We will now look at this point again in more detail with reference once again to Figure 1. The device DAFC comprises a memory MM that will be used to store the first determined residual frequency error f_{el}.
20　Furthermore, the calculation means MC3 will calculate a frequency error F_{el}' defined by formula (II) below:

$$f_{el}' = -\text{sgn}(f_{el})\left(\frac{1}{T} - |f_{el}|\right) \qquad (II)$$

In this formula, sgn denotes the sign function.

Furthermore, the second calculation means MC2
25　formed from an absolute value block followed by a low pass filter will determine the average of the absolute values of a predetermined number of successive residual frequency errors. More precisely, the current average z(k) is defined by the formula (III) below:

$$z(k) = \frac{1}{2}z(k-1) + \frac{1}{2}|x(k)| \qquad\qquad (III)$$

in which k denotes the current iteration corresponding to the current symbol, and x(k) is the current residual frequency error provided by first calculation means

5   MC1.

Comparison means COMP1 can be used to compare the current average output by second calculation means MC2 with a predetermined threshold. If this average is less than the threshold, the comparison means COMP1

10   outputs the value 0. If this average is greater than the threshold, the comparison means COMP1 outputs the value 1.

The result of the comparison made with means COMP1 is examined after a predetermined number of

15   iterations, in other words symbols have been carried out 5 or 6 iterations for example. Although this step is done by software, it is shown schematically in Figure 1 by a register RG containing a constant value equal to 0, and by a switch SW1 that can be switched to

20   input 0 or to input 1 as a function of the value of a control signal output by control means MCTL.

More precisely, as long as the predetermined number of iterations has not been reached, the value of the control signal is equal to zero which puts the

25   switch SW1 to the corresponding input. Consequently, the value zero is output on the control input of a second switch SW2. Switch SW2 is then positioned to its input zero, which puts the output of the filter FLT into relation with the digital-to-analog converter CNA.

30   When the predetermined number of iterations has been reached, the control means MCTL sets the

control signal controlling the switch SW1 to the logical value 1. This moves to its input 1. The value of the control signal of switch SW2 is then fixed by the value output by the comparator COMP1.

5   If this value is equal to 0, corresponding to an average less than the threshold, this means that successive residual frequency errors are converging. Consequently, the output of the filter FLT remains connected to the digital-to-analog converter CNA.

10   However, if the logical value output by the comparator COMP1 is equal to 1, this means that the average of the absolute values of successive frequency errors is greater than the threshold. Therefore, there is no convergence. Consequently, this means that it is

15 very probable that the first stored residual frequency error f_{e1} was not the real frequency error but was its folded version.

      In this case, the switch SW2 is set to its input 1, and the local oscillator VCO is corrected

20 using the residual frequency error f_{e1}' determined by the calculation means MC3. Once this new correction has been applied, the next residual frequency errors drop to a value less than the Nyquist frequency, and the normal cycle starts again with the switch SW2 set to

25 input 0 and the switch SW1 set to input 0. The residual frequency errors then converge towards zero. The memory of filter FLT is then reset to zero before the determined residual frequency error is filtered just after the correction F_{e1}' has been applied.

30   For example, with a DS-CDMA type system with a carrier frequency of 2 GHz and a symbol rate of the controlled signal equal to 15 kHz which leads to a Nyquist frequency of 7.5 kHz, the value output by the

comparator COMP1 after 5 iterations is used and the threshold value is fixed at 500 Hz.

The digital filter FLT is a first order low pass filter. More precisely, the current estimated frequency error $c(k)$ output from the filter (in other words the current digital correction applied to the oscillator) is defined by formula (IV) below:

$$c(k) = (1-b)c(k-1) + bx(k) \qquad \text{(IV)}$$

In this formula, $x(k)$ is the current residual frequency error determined after correction by calculation means MC1. Moreover, b is a coefficient between 0 and 1.

This filter FLT is a matching filter in the sense that the coefficient b is dynamically defined. More precisely, for example, the residual frequency error determined by the calculation means MC1 is compared with a predetermined limiting value, which is shown by a comparator COMP2 in Figure 1. For example, this predetermined limiting value may be equal to the symbol rate divided by 7. If the residual frequency error is greater than this predetermined limiting value, the coefficient b is chosen to be close to 1, for example equal to 1, while the coefficient b is chosen to be close to 0, for example equal to 0 if this residual frequency error is less than the predetermined limiting value. This can give fast convergence of the frequency error towards the target value and avoid oscillations about this target value.

Therefore, the invention can be used to correct residual frequency errors within a correction range equal to the symbol rate. In this respect, note

that if the residual frequency error is close to the absolute value of the Nyquist frequency, the power of the offset Dirac is close to zero, and consequently it is very easy to make a bad estimate of this frequency
5   error. However, this is not a problem for the device according to the invention.

    If the estimate is completely wrong, it is always difficult to have a second estimate error close to the Nyquist frequency, and in this case either one
10   of the two cases mentioned above will be applicable, in other words either a residual frequency error less than the Nyquist frequency, or with an absolute value greater than the Nyquist frequency. Furthermore, any other signal with known content other than the pilot
15   signal can be used for frequency control according to the invention.